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This report discusses the use of the quasi-feedforward linearization technique for improving the range over which an analog fiber-optic system can provide distortionless performance when the optical source is an incoherent device such as an LED. Coherent sources such as a laser are usually dominated by nonlinear effects, such as modal interaction in the fiber and reflections of light from the fiber back into the

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20. ABSTRACT (Continued)

laser cavity, that cannot be eliminated by external compensation. The technique of quasi-feedforward linearization focuses upon deriving a predistorted signal from one optical source that causes distortion cancellation when used to drive a second optical source with similar nonlinear characteristics. The predistorted signal is derived by using two feedforward summations.

A quasi-feedforward compensation system was constructed to measure the bandwidth over which significant distortion cancellation could be achieved using two LEDs having an optical risetime of 14 ns (16 MHz equivalent electrical bandwidth). Over a measurement range of 6 to 18 MHz the improvement in the ratio of signal to second-order intermodulation distortion ranged from 27.5 to 41 dB; the corresponding improvement in second-order linear dynamic range (as defined in the text) was 11.5 to 19 dB. The improvement in the ratio of signal to third-order intermodulation distortion ranged from -7 (negative improvement) to +11 dB; the corresponding modification in third-order linear dynamic range varied from -4 to +2 dB. Absolute values of second-order linear dynamic range measured in 3 kHz noise bandwidth varied from 51 to 55.5 dB before compensation and from 63.5 to 71.5 dB after compensation. Absolute values for third-order linear dynamic range varied from 68.5 to 71 dB before compensation and from 65 to 71.5 dB after compensation.

From the standpoint of performance tradeoff, the small degradation in third-order linear dynamic range at some frequencies is probably a sensible compromise when considering the significant improvement obtained in second-order linear dynamic range from an initially unfavorable value. From the hardware standpoint, space availability, power availability, and temperature control will be critical factors in determining whether the technique is practical for a particular application.

NRL Memorandum Report 5232

Linearization of an Analog Fiber-Optic System Using the Quasi-Feedforward Technique

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December 29, 1983

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LINEARIZATION OF AN ANALOG FIBER-OPTIC SYSTEM USING THE QUASI-FEEDFORWARD TECHNIQUE

INTRODUCTION

To the person concerned with sending a stream of binary data bits over a fiber-optic transmission line, linearity is relatively unimportant provided intersymbol interference is avoided. To the person concerned with transporting multiple radio/communication signals from a receiving antenna to a remote signal processor via a long transmission line, linearity is extremely important. The difference in attitude reflected by these two situations may be reconciled upon consideration of the degree of signal processing occurring before the information is introduced to the transmission line. In the former case the degree of signal processing is high; the information has already been digitized. In the second case it is common that little or no signal processing occurs before the transmission line; the information content from the radio signals is customarily not extracted at the site of the antenna.

This situation poses a dilemma to the system designer who is tasked with interconnecting a remote radio antenna to a signal processor by means of an optical fiber transmission line. On the one hand, the linearity of analog fiber-optic systems may be too poor to merely relay the radio signals over the link by direct intensity modulation. On the other hand, digitization of the radio signals at the site of the antenna may be tantamount to moving the signal processor to the forward end of the transmission line -- a measure that will rescue the fiber-optics but may destroy the intended system architecture. Consequently the fiber-optic system designer looks for a compromise or alternative approach that permits retaining most of the complex signal processing equipment at the far end of the link while simultaneously overcoming the linearity problem.

Reference [1] suggests some link design techniques that make the best use of limited linear dynamic range margin and provides quantitative limits on the linear dynamic range that can be achieved with LED and laser sources. Specifically, the maximum second- and third-order linear dynamic range (LDR_2 & LDR_3) that can be obtained in 3 kHz bandwidth with a 500 meter link length are shown to be:

- Typical multi-longitudinal mode laser

  LDR_3: 62 dB

  LDR_2: 50 to 55 dB

- Super-radiant laser (50 to 70 longitudinal modes)

  LDR_3: 68 to 69 dB

  LDR_2: 50 dB

- LED

  LDR_3: 70 to 75 dB

  LDR_2: 53 to 58 dB

Manuscript approved October 4, 1983.

Here linear dynamic range is defined to be the difference in decibels between the maximum signal (unity distortion-to-noise ratio for the particular order of distortion considered) and the minimum signal (unity signal-to-noise ratio) for a specified noise bandwidth.

In reference [2] it was shown that the use of an intermediate electrical FM carrier to intensity modulate the optical source will:

- Result in large demodulated signal-to-noise ratios (SNRs) in the presence of small carrier-to-noise ratios (CNRs).

- Provide superior overall analog performance for multi-carrier frequency-division-multiplexing (FDM) systems particularly in the presence of large link loss.

This report addresses the prospect of making an LED source function more linearly (i.e. with less distortion) by an external predistortion technique known as quasi-feedforward compensation. The modifier "quasi" is used because although standard feedforward techniques are employed, performance improvement is achieved by predistorting an LED by the distortion derived from another similar LED. In true feedforward linearization of a device such as a power amplifier, linearization is achieved by injection of inverted-phase distortion components into the output of the amplifier, but in this case the distortion is derived from the amplifier itself -- not a duplicate unit.

The method of achieving quasi-feedforward linearization is shown schematically in Figure 1. The outputs of two signal generators operating at slightly different frequencies are combined to form a clean, two-tone interrogating signal (not part of the linearizing circuit). The interrogating signal (or any signal input to the system) is divided into two parallel branches that are, in principle, exactly matched in amplitude and phase. One branch is used to modulate an LED that has a very short optical link to a photodetector.

We assume that the LED is nonlinear but that the photodetector and associated amplifiers are perfectly linear. All electro-optical components of this branch must have a flat frequency response and a linear phase vs. frequency response over the band in which quasi-feedforward linearization is attempted; otherwise, the optical branch cannot be duplicated by an equivalent parallel electrical branch with frequency-independent delay and attenuation.

The first power combiner is chosen to have one inverting input (for the "optical" branch) and one noninverting input for the parallel electrical branch. The parallel electrical branch is assumed to be ideal in that the signal arrives at the power combiner at the same amplitude and phase as does the signal arriving via the optical branch. The output of the power combiner consists, in principle, only of the inverted-phase distortion products associated with the nonlinearity of the LED. In practice there will also be residuals from a less-than-perfect cancellation of fundamental signals. The cancellation will be broadband to the extent that the frequency dependent

gains and phase shifts of the two parallel branches are indentical. The two branches can be checked for similarity of phase/magnitude characteristics in the laboratory by temporarily substituting a network analyzer at the position of the first combiner.

The purpose of the second combiner is to restore the signal to the inverted-phase distortion components, thereby deriving the predistorted signal to drive the second similar LED that is used to modulate the optical link. Proper setup of the second power combiner is achieved when the length (delay) and loss of the input cables are adjusted to provide the same magnitude but inverted phase relationships between the signal and distortion products as those existing at the output of the photodetector.

A network analyzer cannot be used directly to eliminate the delay differential in the branches to the second combiner because the two branches carry different signals. One branch carries pure signal; the other carries inverted-phase distortion products and a small residual of inverted-phase signal. However, if the electrical reference to the first combiner is temporarily removed (and that input is properly null-terminated), then the second combiner will function as a signal canceler. The quality of the cancellation in the second combiner can be observed on a spectrum analyzer while the signal generators are swept in frequency.

Final adjustments are made by observing the residual distortion at the output of the second or "link" photodetector. This process is difficult because it must always be verified that any adjustment that reduces distortion at a single testpoint also reduces distortion over the entire bandwidth of interest. It is extremely easy to make adjustments that make significant improvements in a narrow frequency band while simultaneously degrading performance outside that band.

There are two subtleties that must be heeded if the circuit is to be made to work properly. Firstly, the amplifiers between the second combiner and LED#2 must be noninverting overall. This requirement is best explained as follows. Assume that the signal introduced to the feedforward system is a finite, positive, unidirectional excursion. This signal will drive LED#1 along its transfer characteristic exclusively above the bias point. If the derived predistortion signal drives LED#2 in the opposite sense, the distortion canceling scheme will obviously not work.

The second important point concerns crosstalk between the ports of the power combiner. Typically, broadband couplers/splitters will not provide better than 35 dB isolation between ports. When distortion cancellation as large as 40 dB is realized with the feedforward circuity, feedthrough from inadequate coupler-port to coupler-port isolation can be exceedingly detrimental. Our system was very sensitive to leakage of distortion from the optically derived input into the second combiner. This leakage occurred as a result of crosstalk between the 180° and 0° ports of the first combiner followed by a low-loss path to the second combiner. (The attenuators shown in this pathway in Figure 1 were not present until the problem was recognized.) Fortunately, direct passage through the combiners of the port-to-port crosstalk was not a serious problem.

To achieve a given degree of signal component power reduction in a feedforward cancellation, specific tolerances must be met in terms of amplitude and phase mismatch. Assume that a signal (or distortion) component specified by $y_1 = A\sin\omega t$ is to be partially eliminated by subtraction with a signal of slightly different amplitude and phase $y_2 = (A + \varepsilon)\sin(\omega t + \Delta)$. The instantaneous cancellation is given by $y_c = y_1 - y_2$, but the physical quantity of interest is the mean square value of y_c over one cyclic period,

$$\langle y_c^2 \rangle = A^2 + \varepsilon A + \varepsilon^2/2 - A(A + \varepsilon)\cos\Delta \tag{1}$$

The fractional power cancellation is given by

$$\frac{\langle y_c^2 \rangle}{\langle y_1^2 \rangle} = 2 + 2\,\varepsilon/A + \varepsilon^2/A^2 - 2(\cos\Delta)(A + \varepsilon)/A \tag{2}$$

Contours of constant power cancellation are shown in Figure 2 as a function of amplitude and phase mismatch. This illustration provides excellent insight into the difficulty in constructing circuitry that will provide large distortion cancellation. To achieve 20 dB cancellation, for example, the two waveforms must be matched in amplitude to with 0.83 dB assuming zero phase mismatch. If the same degree of cancellation is to be achieved over a given bandwidth, both branches of the feedforward circuit must exhibit a differential gain less than 0.83 dB over the entire bandwidth. To achieve 30 dB cancellation we must satisfy the very stringent requirements of less than 0.3 dB differential gain and 2 degrees differential phase.

A final issue of importance is the relationship between distortion reduction and the corresponding increase in linear dynamic range. Consistent with our previous verbal definition of linear dynamic range we may write (in units of dB)

$$LDR_i = S - N - DNR_i/i \tag{3}$$

where DNR_i is the distortion-to-noise ratio observed from the i^{th} order of distortion at a signal level S and a noise level N. For example, if second-order intermodulation distortion is being considered and we observe S = 10 dBm, N = -90 dBm, and DNR = 20 dB, then $LDR_2 = 10 + 90 - 20/2 = 70$ dB.

If the DNR is reduced by I dB without changing the signal-to-noise ratio, then

$$LDR_i = S - N - (DNR_i - I)/i \tag{4}$$

where the DNR was the distortion-to-noise ratio measured prior to the improvement I. Consequently, for a given cancellation of I dB, second-order linear dynamic range increases by I/2 dB and third-order linear dynamic range increases by only I/3 dB. Consequently, reports of feedforward cancellation appear more spectacular when reported in terms of distortion cancellation than when reported in terms of linear dynamic range improvement, particularly for high orders of distortion.

The pioneering work in quasi-feedforward linearization was done by Straus, Patterson, Blenman, and Witkowicz of Bell Northern Research, Ltd. [3, 4]. Their work indicated that second-order distortion cancellation up to 35 dB and third-order distortion cancellation up to 20 dB could be achieved with a fiber-optic system. Unfortunately, their work does not provide information regarding the bandwidth over which the specified cancellation can be achieved nor are specific cancellation versus frequency data presented. In order to develop perspective on the difficulty in achieving a large cancellation magnitude-bandwidth product, a research program was begun at NRL in anticipation of a technology payoff for a towed communication buoy application.

SYSTEM IMPLEMENTATION

The apparatus used in the quasi-feedforward experiments is shown in Figure 3. The tuned bandpass filters used to clean up the CW tones from the signal generator were laboratory-constructed devices. The summation of the tones was actually accomplished by internal inductive coupling rather than by a simple T-junction as shown.

The purity of the combined test tones is demonstrated in Figures 4, 5, and 6. The SNR of the summed tones (6.0 and 6.1 MHz) at the output of the tuned bandpass filters is shown in Figure 4. The rise in the noise floor in the vicinity of the tones is caused by the phase noise of the signal generators. The lower noise plateau is the noise floor of the spectrum analyzer with 20 dB of RF attenuation. If noise is measured at least 200 kHz away from the center of the test tones, SNRs of at least 80 dB in 3 kHz analysis bandwidth are achieved.

The test tones are shown again in Figure 5 with 30 dB RF attenuation (to diminish distortion originating from spectrum analyzer overload), 50 kHz per division horizontal sensitivity, and 300 Hz analysis bandwidth. The $2f_1 - f_2$ and $2f_2 - f_1$, third-order intermdulation products that occur at 5.9 and 6.2 MHz are below system noise and are, consequently, less than -86 dBc (dB below carrier).

In Figure 6 the frequency span has been increased to allow observation of second harmonic, second order IM(f_1+f_2), and third harmonic distortion. Third harmonic distortion is barely perceptible above the noise at - 79 dBc. Second harmonic distortion is not visible, but the $f_1 + f_2$ second-order IM, which is normally 6 dB above second harmonic products, is visible (emphasized by the dot marker) at -77 dBc.

To split the test signal into two parallel branches with small frequency dependent differential gain and phase shift, a four-port ANZAC power splitter was used because of its availability. The two unused ports were null-terminated with coaxial in-line 50 ohm loads.

The LEDs used were Laser Diodes, Inc. type IRE 160FA. These devices were known from previous measurements [1] to have a third-order linear dynamic range LDR_3 between 70 and 75 dB and a second-order linear dynamic range LDR_2 of about 51 to 59 dB in a 3 kHz bandwidth. The device risetime is about 14 ns corresponding to an electrical modulation bandwidth of 16 MHz. The AC drivers used to modulate the LEDs are shown in Figures 7 and 8. Simple RC coupling circuits with nominal 50 ohm input impedance were used. The more intricate RC structure for the second LED driver was required to equalize the frequency response for both sources, probably because of a slightly different optical risetime for each LED. The 16 MHz bandwidth of the LED will limit the effectiveness of the feedforward linearization scheme at high frequencies because it is not possible to duplicate low-pass filter characteristics near the 3 dB point by a fixed delay line.

The optical detectors are shown schematically in Figures 9 and 10. Both circuits use a MERET R-1900 Hybrid optical detector that combines a PIN photodiode with a transimpedance amplifier in a TO-5 package. The external components added to the MERET package differ only in that better power supply decoupling and high frequency compensation were used with the first optical detector because of its very critical location in the feedforward circuit.

Referring again to Figure 3, the delay associated with the optical branch is duplicated in the parallel electrical branch with a fixed length of RG-58 coaxial cable that was trimmed to exact length using multiple BNC coaxial adapters and an adjustable air delay line. The LORCH 266B power combiner inverts the phase of the "optical" branch relative to the parallel electrical branch to achieve a subtraction of the fundamental signal, leaving a residual of inverted-phase distortion products at the output. The degree of fundamental cancellation over the 2 to 32 MHz frequency range is shown in Figure 11. More than 35 dB of cancellation was achieved over much of this range. At the higher frequencies the limited bandwidth of the LED is responsible for the diminished cancellation. At the lower frequencies the .022 microfarad coupling capacitor in the LED driver circuit is probably responsible for the diminished cancellation; based on 50 ohm load resistance, the .022 microfarad capacitor will produce a 4.14° phase shift at 2 MHz which will limit the cancellation potential to 22.8 dB.

The LORCH 261B coupler is used as a power combiner to add the signal to the inverted phase distortion components--thus completing the formation of the predistorted signal to drive the second LED. Fifty-ohm loads were used at the input to the couplers to achieve proper impedance matching. Coaxial in-line attenuators were used to achieve proper amplitude balance. The 6 dB attenuator that is T-ed into the 3 dB attenuator on one side but unterminated on the other side was used to provide a small amount of additional loading required to achieve the proper amplitude balance.

The performance of the second power combiner (LORCH 261B) and associated feeder lines was verified by observing the output on a spectrum analyzer when the electrical reference branch to the first power combiner was temporarily removed. With this configuration the second combiner becomes a signal canceler. The cancellation achieved from 2 to 30 MHz is shown in Figure 12 (dominant trace). The fainter trace shows the cancellation at the first combiner and is used as a reference for assessing the performance of the second combiner. (The slightly different appearance in Figures 11 and 12 of the cancellation trace from the first combiner resulted from some additional circuit fine adjustments that were implemented between the two sets of measurements.)

The series of amplifiers between the second LED and the LORCH 261B combiner is used to boost the signal to a level that provides maximum distortion cancellation. This amplitude level will be approximately the same as that used to drive the first LED. The LORCH 266B coupler was used as an inverter to eliminate the signal inversion resulting from the cascading of the three boosting amplifiers. We have explained, in the INTRODUCTION, the necessity to drive both LEDs in the same directional sense relative to the transfer characteristic in order to achieve distortion cancellation.

The optical transfer characteristics for the two LEDs are shown in Figure 13. The difference in slopes of these curves is a matter of interest. It is not necessary for the slopes of the two transfer characteristics to be identical to achieve good distortion cancellation; however, it is important that the relative curvature of the transfer characteristics be nearly identical. For example let us assume that the transfer characteristic can be written mathematically as

$$P = a(b_1 I + b_2 I^2 + b_3 I^3) \tag{5}$$

where P is the output optical power, I is the drive current, the b's are the coefficients of a third-order polynomial describing the transfer characteristic, and "a" is a scaling multiplier. Assuming two-tone modulation we write

$$I = A\cos\alpha t + B\cos\beta t \tag{6}$$

The output optical power at the fundamental angular frequencies α and β is

$$\begin{aligned} &a(b_1 A\cos\alpha t + b_1 B\cos\beta t) \\ &+.75ab_3 A(A^2 + 2B^2)\cos\alpha t \\ &+.75ab_3 B(B^2 + 2A^2)\cos\beta t \end{aligned} \tag{7}$$

where the last two terms are small additions to the intended fundamental output (first term) caused by the third-order polynomial term. Second-order distortion is given as

$$0.5ab_2(A^2\cos 2\alpha t + B^2\cos 2\beta t)$$
$$+ ab_2AB(\cos(\alpha+\beta)t + \cos(\alpha-\beta)t) \qquad (8)$$

and third-order distortion as

$$0.25ab_3(A^3\cos 3\alpha t + B^3\cos 3\beta t)$$
$$+ 0.75ab_3A^2B(\cos(2\alpha+\beta)t + \cos(2\alpha-\beta)t) \qquad (9)$$
$$+ 0.75ab_3B^2A(\cos(2\beta+\alpha)t + \cos(2\beta-\alpha)t)$$

Since the ratio of the desired fundamental signal to any distortion product is independent of "a", the signal-to-distortion ratio will be constant for any transfer characteristic that has polynomial coefficients that differ only by a constant multiplier and that is driven by carriers of the same amplitude.

The important consequence of the different slopes of the transfer characteristics for the two LEDs is reduction in SNR. If the second LED is modulated with the same amplitude carrier(s) as used for the first in order to achieve the same signal-to-distortion ratio (assuming similar relative curvature of the two transfer characteristics), the ratio of the peak-to-peak swing in optical power will equal the ratio of the differential slopes at the bias point. For our experiments both LEDs were biased for 105 microwatts of average optical power; at this bias level the ratio of the smaller to larger slope is 0.547 which leads to a 5.2 dB reduction in electrical SNR (20 log (slope ratio)) at the output of the optical detector. Given the situation that both transfer characteristics must be modulated to produce the same signal-to-distortion ratio with a 5.2 dB difference in signal-to-noise ratio, the second LED, operating by itself will be capable of developing an LDR_3 that is 5.2(2/3) = 3.5 dB less than that of the first LED and an LDR_2 that is 5.2(0.5) = 2.7 dB less than that of the first LED. In summary, the quasi-feedforward technique is potentially more disadvantaged by the smaller (slope-induced) sensitivity of the second LED than by an inability to achieve distortion cancellation.

In order to convey some perspective on the delays associated with the various branches of the quasi-feedforward circuit, the lengths of coaxial cables used from point-to-point are given in Table 1. This information will also serve as a starting point for future investigations.

Table 1 — Critical Transmission Line Lengths

FROM	TO	LENGTH
4-H50-4 SPLITTER	50 Ω TERM OF LED #1	50.8 cm
LED #1 (OPTICAL FIBER)	OPTICAL DETECTOR #1	110.5 cm
OPTICAL DETECTOR #1	AIKEN 1003-2 AMPL	17.8 cm
AIKEN 1003-2 AMPL	50 Ω TERM TO LORCH 266B	182.8 cm
4-H50-4 SPLITTER	WEINSCHEL ATTN	12.7 cm
WEINSCHEL ATTN	GR-874 DELAY LINE	50.8 cm
INPUT GR-874	OUTPUT GR-874	90.2 cm
GR-874	INPUT ATTN OF LORCH 266B	4.88 cm
50 Ω INPUT TERM TO LORCH 266B (0°)	INPUT ATTN OF LORCH 261B	50.8 cm
OUTPUT LORCH 266B	50 Ω TERM OF LORCH 261B	3.8 cm

RESULTS

Measurements of signal, distortion, and noise levels were made at the output of the first LED/optical detector combination and at the output of the linearized system using an HP 8568A spectrum analyzer. Two-tone, equal amplitude excitation with frequency separation of 100 kHz was used for all measurements over a range of 6 to 18 MHz. The frequency range over which measurements were taken was limited primarily by the range of the bandpass filters used to clean-up the output of the signal generators. Measurements beyond 18 MHz would probably not have been rewarding because of the deterioration in performance already evident at 18 MHz in the data that will be presented below. Diminished performance at the high frequency extreme of the measurements is attributed to the nominal 16 MHz bandwidth of the LED which introduces differential phase shifts that cannot be matched by a fixed-delay, parallel feedforward circuit branch. Performance below 6 MHz is also expected to deteriorate because of phase shift introduced by the coupling capacitors to the LED drivers.

The particular distortion products that were monitored were f_1+f_2 second-order IM (6 dB larger than second harmonic products), $2f_1-f_2$ third-order IM (occurring 100 kHz below f_1), and $2f_2-f_1$ third-order IM (occurring 100 kHz above f_2). Since LEDs produce significantly more second-order distortion than third-order distortion, we should anticipate that the greatest improvement achieved with quasi-feedforward linearization will be with the second-order products.

The improvement in signal to second-order distortion S/IM_2 and signal to third-order distortion S/IM_3 ratios are shown as a function of frequency in Figure 14. As previously implied, the improved performance of the linearized system was gauged relative to the performance of the first (non-linearized) LED. A minimum of 27 dB of improvement in S/IM_2 was achieved over the entire 6 to 18 MHz band with improvements as large as 40 dB in the 13 to 14 MHz range. Approximately 10 dB improvement in S/IM_3 was achieved from 6 to 12 MHz, but thereafter the improvement diminishes with performance actually being worse (negative improvement) beyond 14 MHz.

As discussed in the INTRODUCTION, the most spectacular presentation of the linearization improvement data is in terms of S/IM improvement as shown in Figure 14. The improvement in linear dynamic range, as defined in the INTRODUCTION, is only about one-half of the S/IM_2 improvement for second-order linear dynamic range and only about one-third of the S/IM_3 improvement for third-order linear dynamic range. The improvement in linear dynamic range is shown in Figure 15 as a function of frequency. The improvement in LDR_2 exceeds 10 dB over the entire test range of 6 to 18 MHz and achieves a maximum of 19 dB at 14 MHz. The improvement in LDR_3 is less than 3 dB from 6 to 12.5 MHz and actually deteriorates by as much as 4 dB between 12.5 and 18 MHz. It should be emphasized that the LDR improvements shown in Figure 15 were determined by subtracting the LDR values determined at each frequency for the linearized and nonlinearized systems rather than simply inferring the LDR improvement from the S/IM improvement data according to the formalism developed in the INTRODUCTION which assumes identical signal to noise ratios before and after the cancellation process. This procedure resulted in LDR improvements that were a few dB smaller than would have been estimated exclusively on the basis of S/IM improvement.

In Figure 16 the absolute LDR_2 in 3 kHz noise bandwidth is shown as a function of frequency for both the linearized and noncompensated systems. Without quasi-feedforward linearization the LDR_2 spans a range of 51 to 55.5 dB. After linearization the LDR_2 spans a range of 63.5 to 71.5 dB depending upon frequency.

In Figure 17 the absolute LDR_3 is shown before and after linearization. Before compensation the LDR_3 spans the range of 68.5 to 71 dB; after compensation the LDR_3 spans the range of 65 to 71.5 dB. Considering the dramatic improvement in LDR_2 from an initially unfavorable value, the slight degradation in LDR_3 at some frequencies is probably a worthwhile compromise.

CONCLUSIONS

The quasi-feedforward linearization technique was used to reduce distortion in an LED with a nominal electrical modulation bandwidth of 16 MHz. Over a measurement range of 6 to 18 MHz the improvement in signal to second-order intermodulation distortion ranged from 27.5 to 41 dB; the corresponding improvement in second-order linear dynamic range was 11.5 to 19 dB. The improvement in signal to third-order intermodulation distortion ranged from -7 (negative improvement) to + 11 dB; the corresponding modification in third-order linear dynamic range varied from -4 to +2 dB. Absolute values of second-order linear dynamic range measured in 3 kHz noise bandwidth varied from 51 to 55.5 dB before compensation and from 63.5 to 71.5 dB after compensation. Absolute values of third-order linear dynamic range varied from 68.5 to 71 dB before compensation and from 65 to 71.5 dB after compensation.

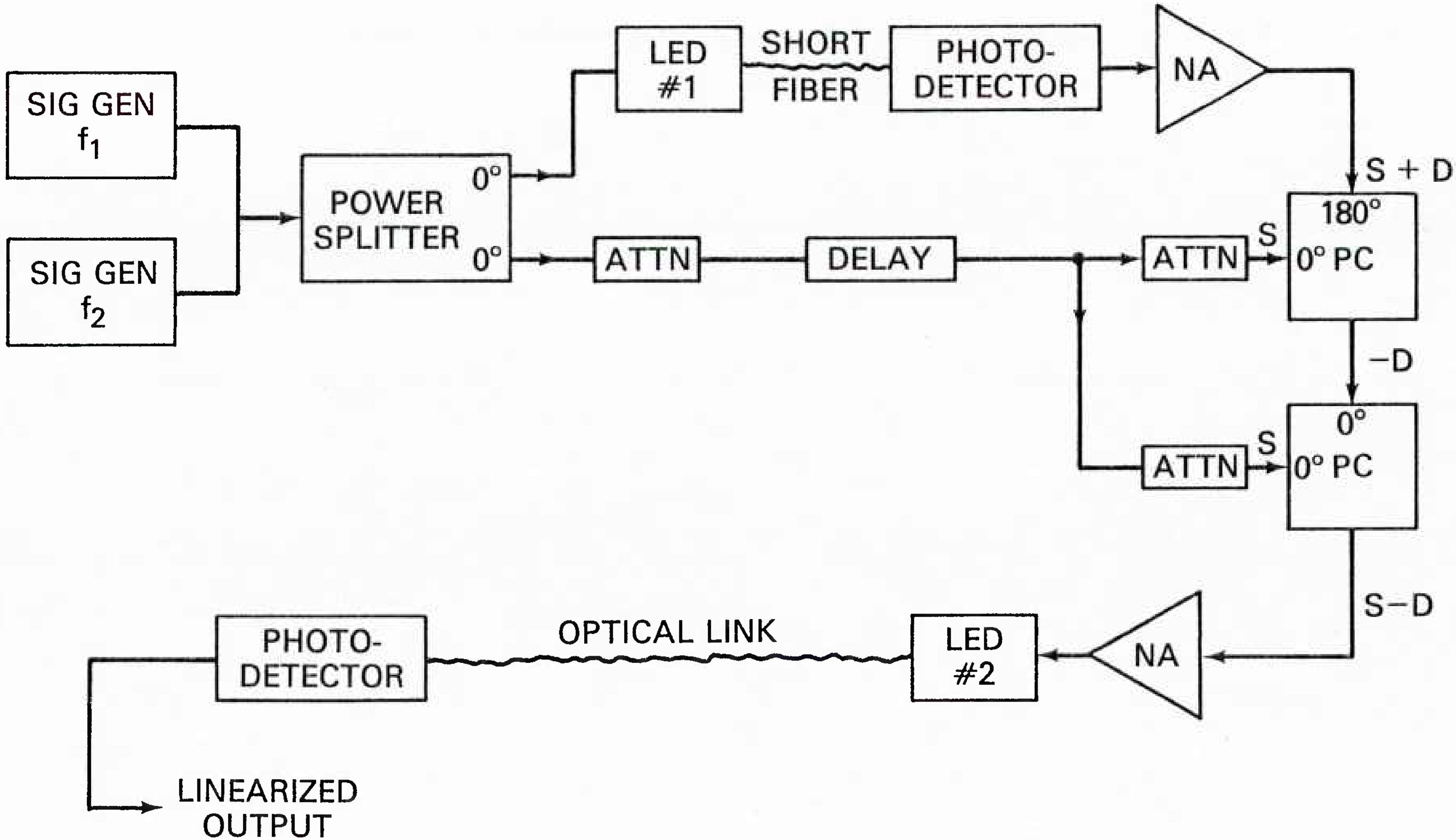
From the standpoint of performance tradeoff, the small degradation in third-order linear dynamic range at some frequencies is probably a sensible compromise when considering the significant improvement obtained in second-order linear dynamic range from an initially unfavorable value. From the hardware standpoint, it is doubtful whether the quasi-feedforward technique will be beneficial for all applications. The major stumbling blocks to implementing the technique are:

- Amplifier requirements. The output of the first optical detector must be boosted to a level in the range of -5 to 0 dBm with all distortion products less than -75 dBc to drive the second LED. This requires medium power amplifiers with relatively large second- and third-order intercept points.

- Differential gain and phase stability. All circuit branches to the first and second power combiners must retain close tolerances on gain and phase shift over a wide frequency range. To obtain 20 dB cancellation at a combiner, relative phase shift must be 5.7 degrees or less with zero amplitude mismatch and amplitude match must be less than 0.83 dB for zero phase mismatch.

The implementation of quasi-feedforward linearization appears to be profitable and practical in an environment where temperature can be controlled and sufficient space and power is available for the additional components. For remote platform applications such as towed buoys and arrays and tethered kites and balloons, the practicality of the technique is doubtful. Of particular concern is the necessity to retain close tolerances on differential gain and phase in an environment in which temperature cannot be controlled.

We have had very limited experience with the temporal and thermal stability of the quasi-feedforward apparatus. The maximum temperature difference that occurred during laboratory operation of the device was about 8 C, and no detrimental effects were observed over this temperature range. Probably the longest period over which performance was monitored without some change being made was one week; performance was stable over this period.



PC: POWER COMBINER
NA: NON-INVERTING AMPLIFIER

Fig. 1 — Illustration of method used to achieve quasi-feedforward linearization.

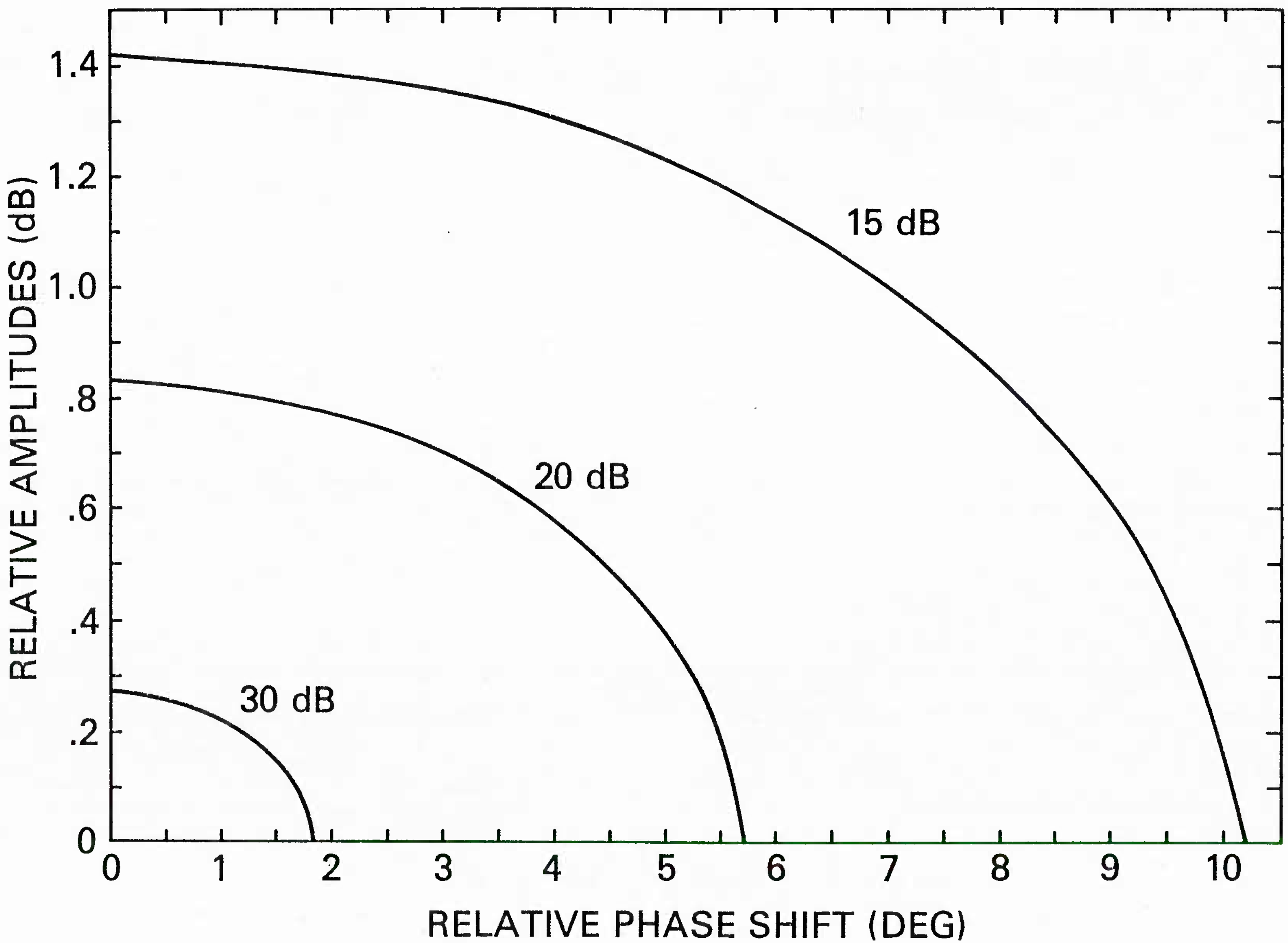


Fig. 2 — Contours of constant cancellation versus relative phase and amplitude.

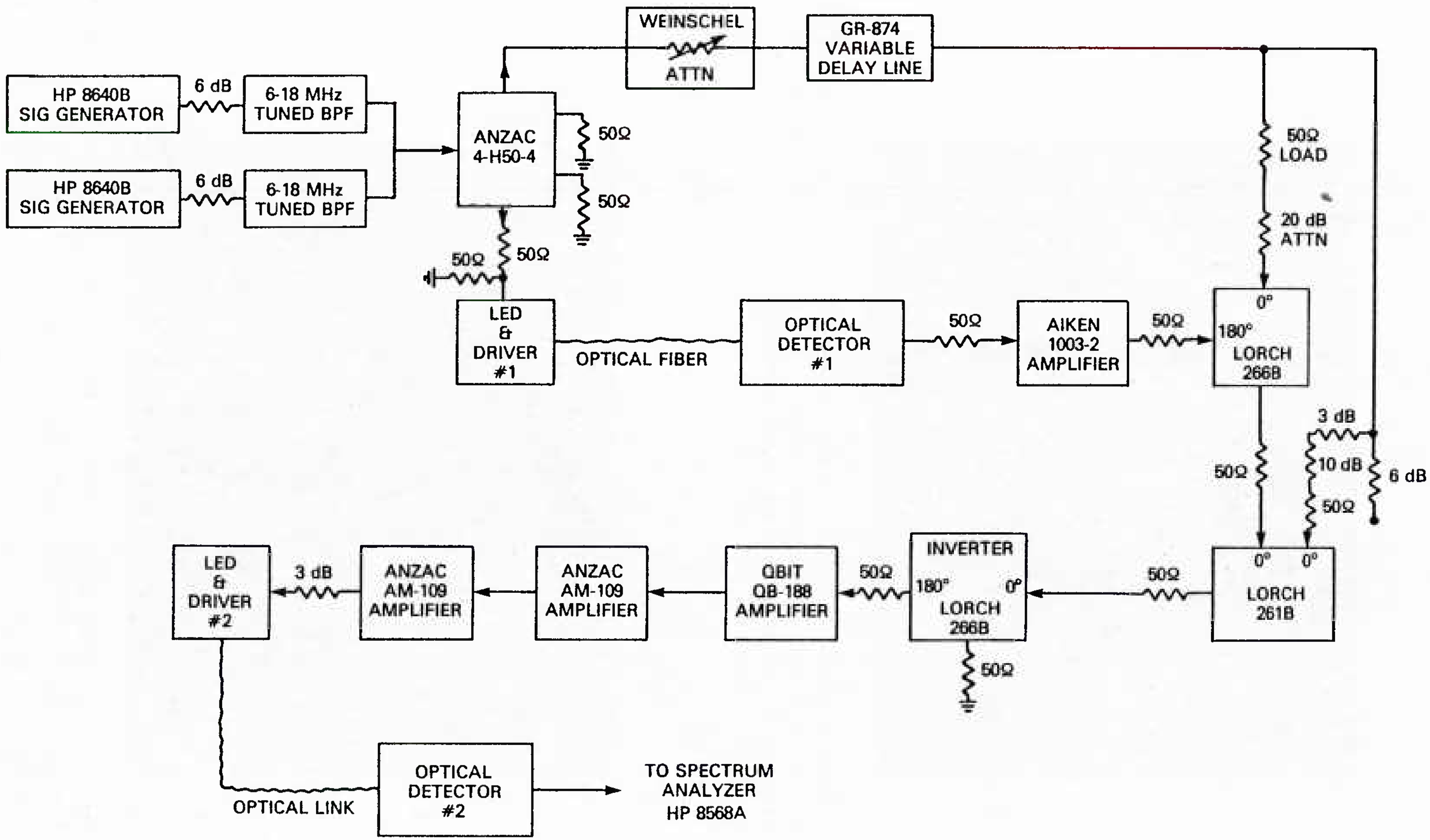


Fig. 3 — Block diagram of specific components used to construct the quasi-feedforward linearization system.

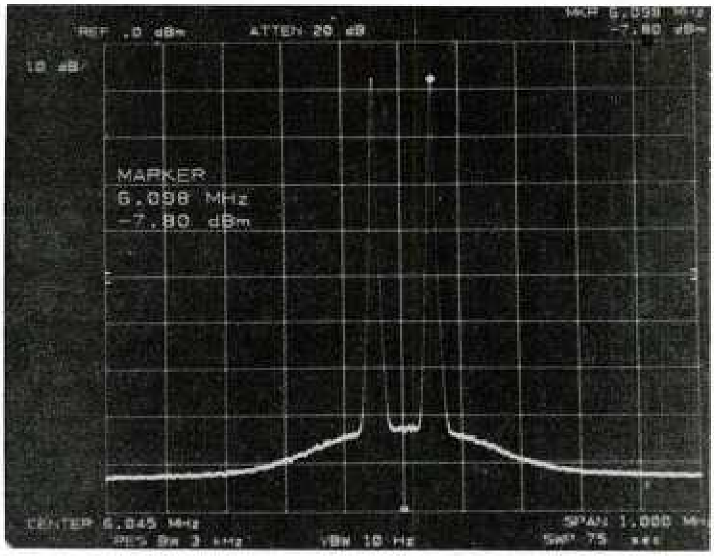


Fig. 4 — Frequency spectrum showing signal-to-noise ratio of dual-tone interrogating signal at the output of the tuned bandpass filters. (The rise in the noise floor near the signals is caused by the phase noise of the signal generator.)

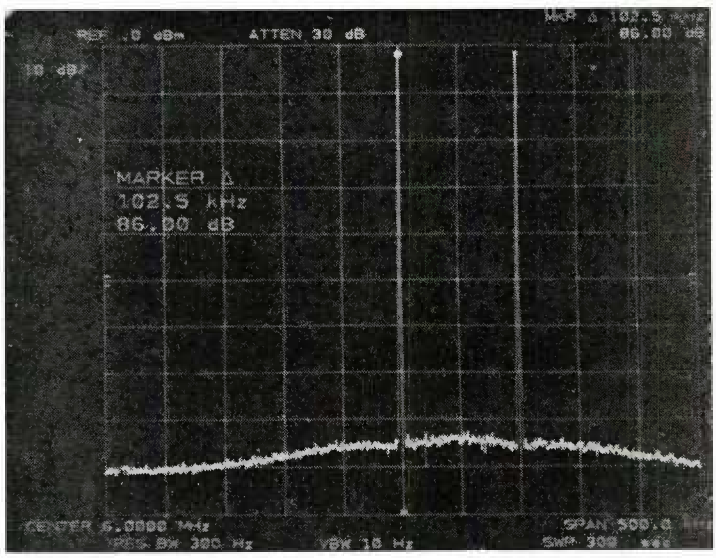


Fig. 5 — Frequency spectrum showing that the third-order intermodulation products $2f_1$-f_2 and $2f_2$-f_1 of the interrogating signal are less than - 86 dBc.

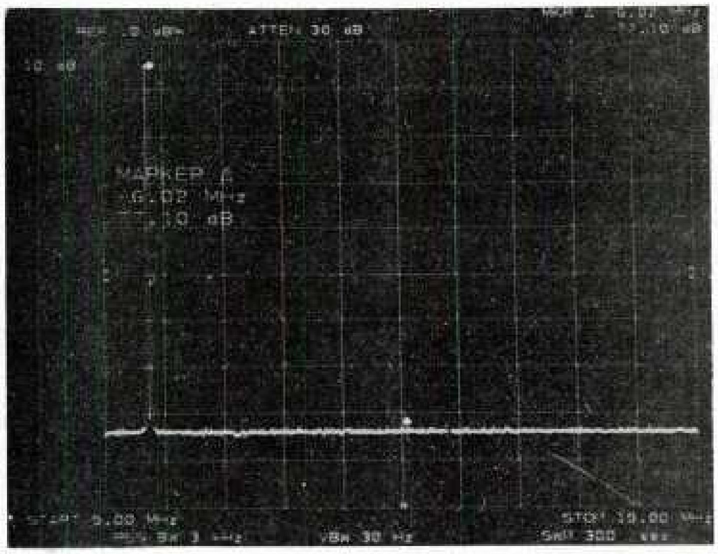


Fig. 6 — Frequency spectrum showing that the second-order intermodulation product f_1+f_2 of the interrogating signal is at - 77 dBc.

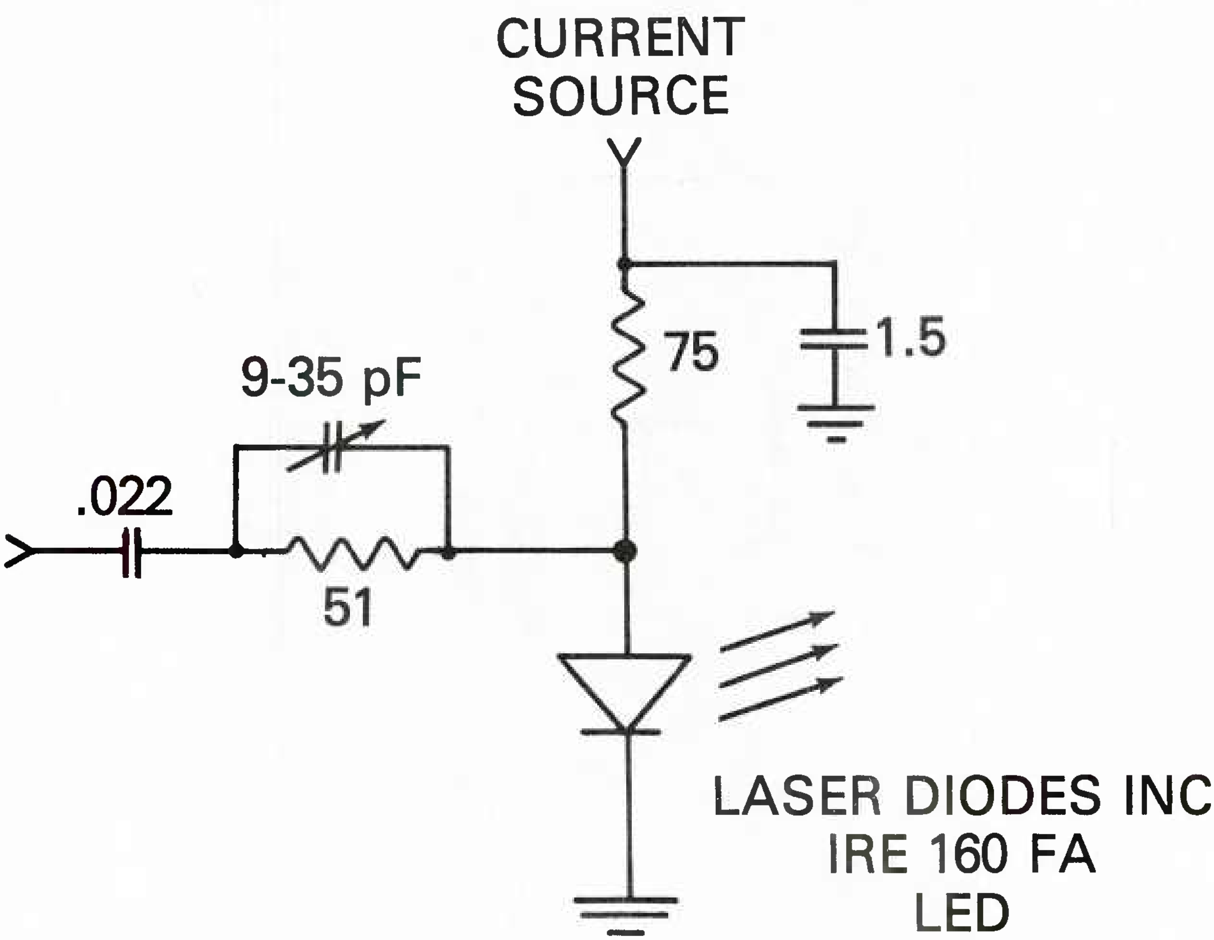


Fig. 7 — Circuit components for biasing and modulating the first LED.

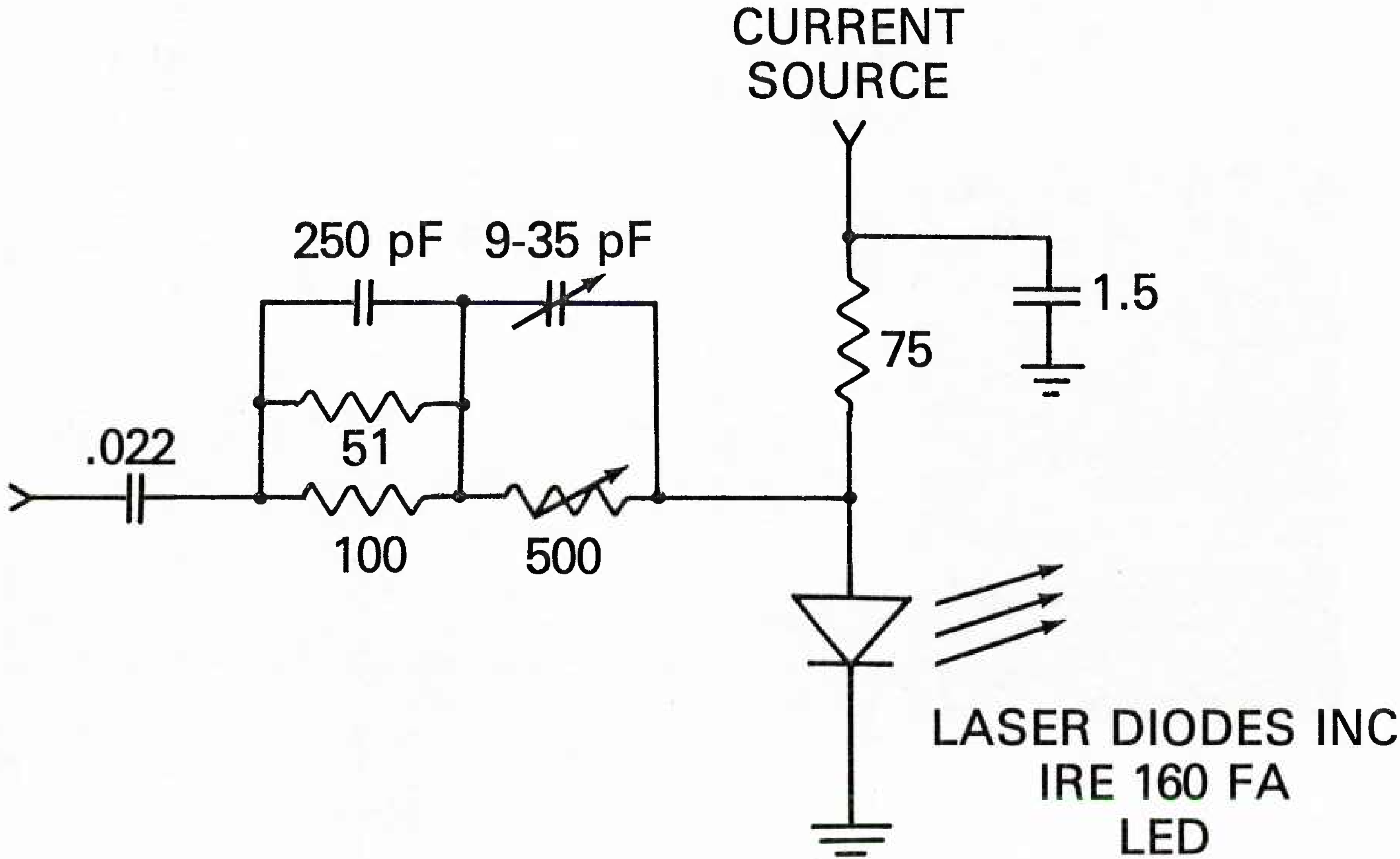


Fig. 8 — Circuit components for biasing and modulating the second LED.

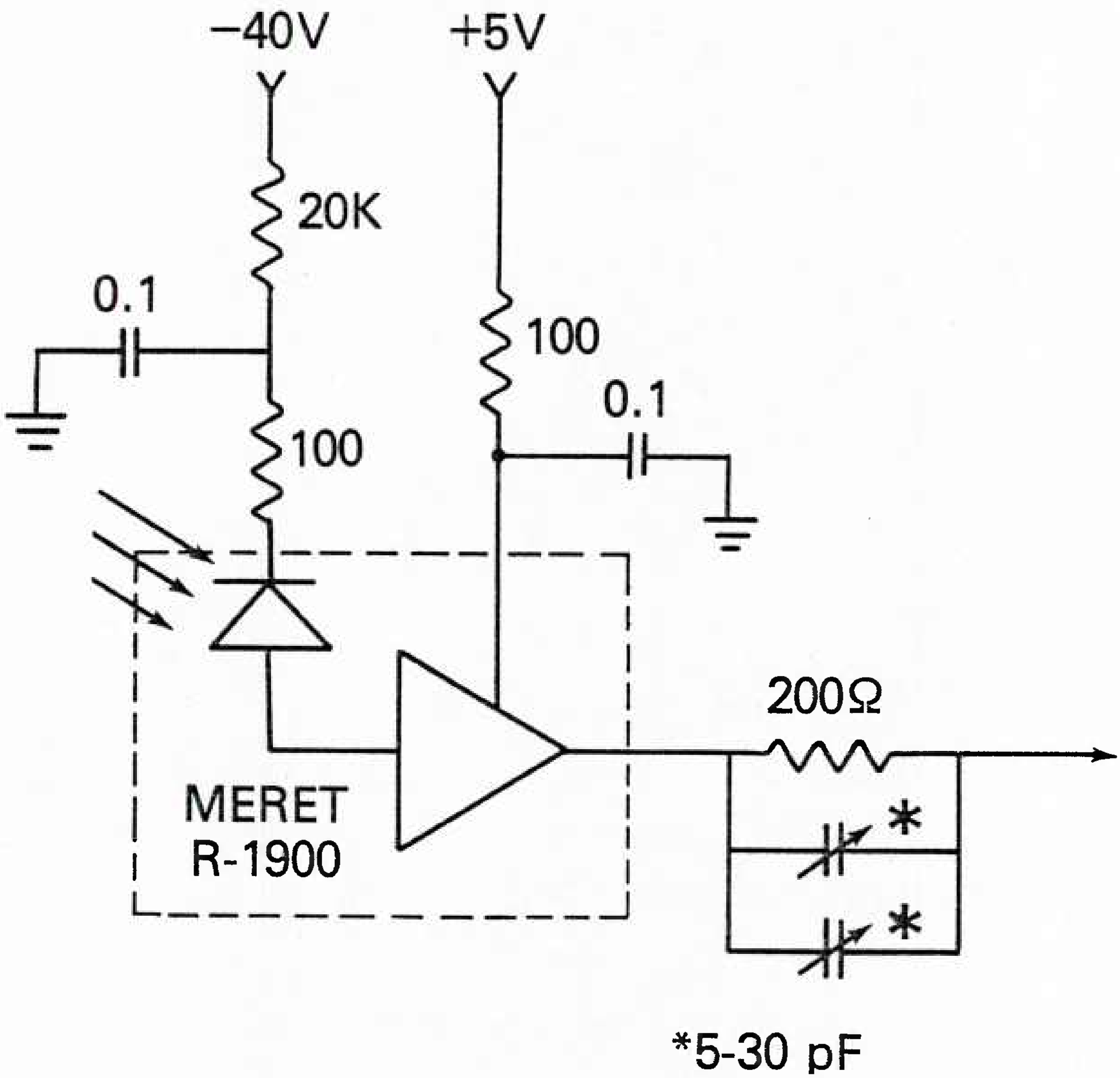


Fig. 9 — Circuit components for the first optical detector.

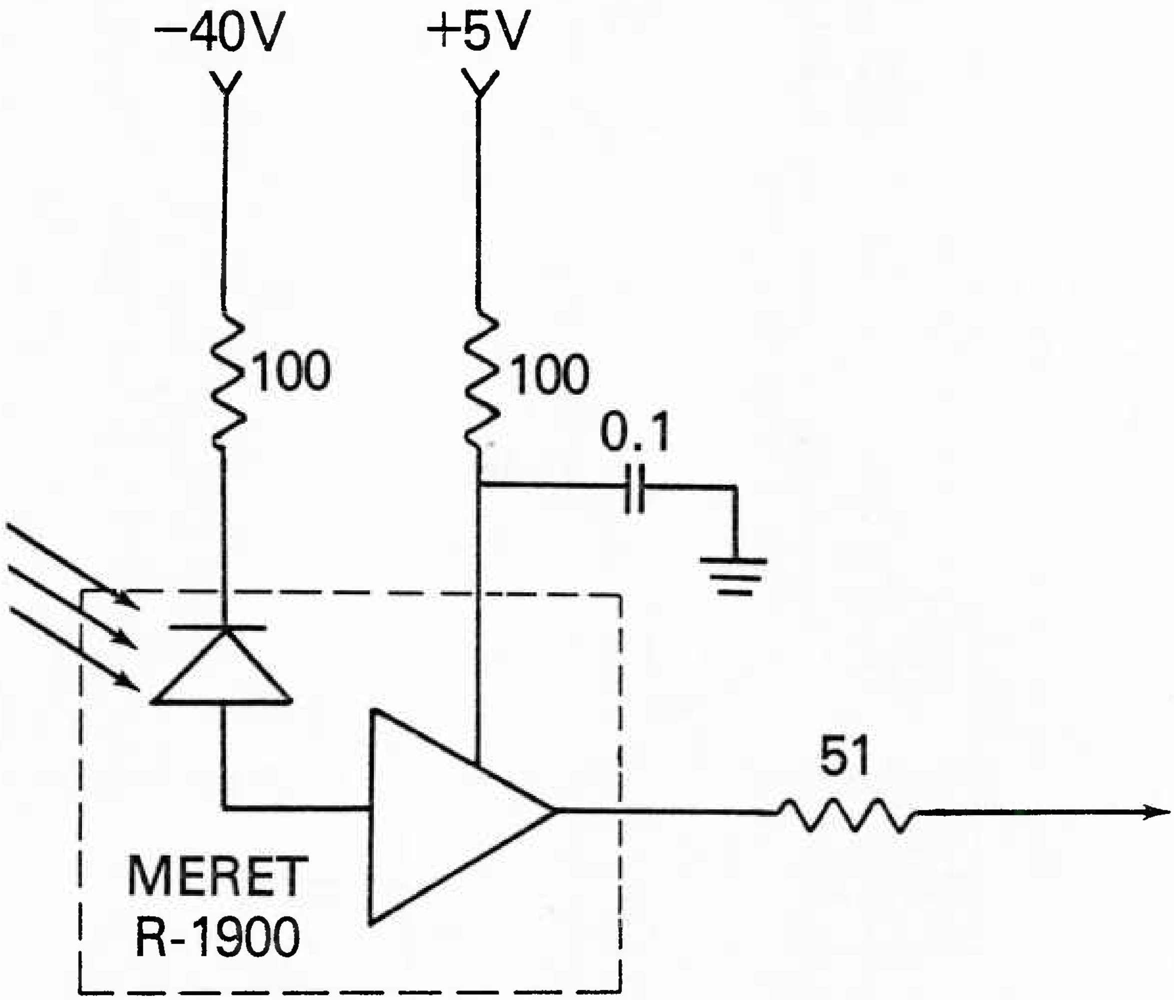


Fig. 10 — Circuit components for the second optical detector.

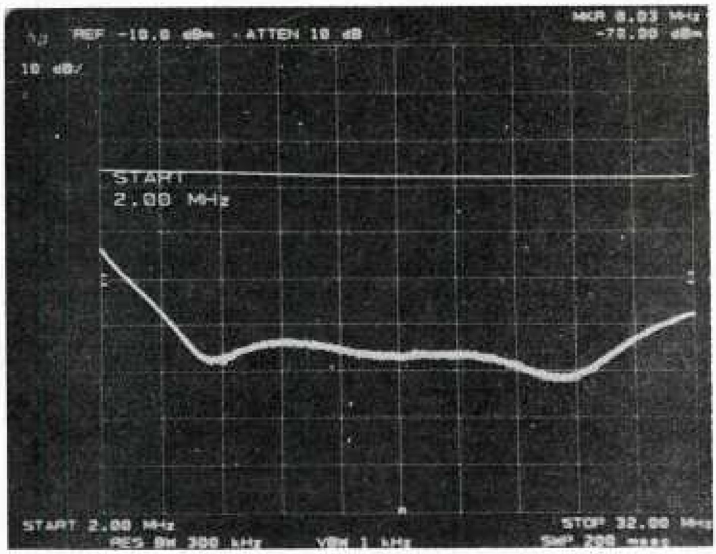


Fig. 11 — Frequency dependence of the cancellation achieved at the first power combiner. Upper line is the signal reference level; lower curve is the residual from the output of the power combiner.

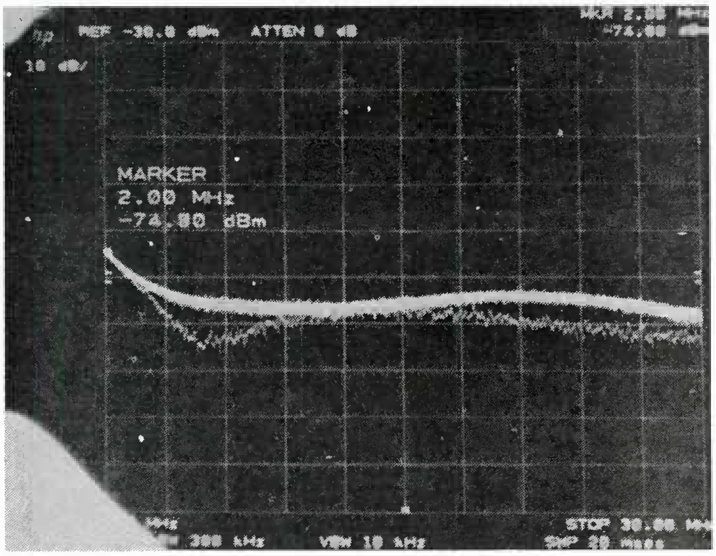


Fig. 12 — Residual from output of second power combiner when electrical reference line to first power combiner is temporarily removed (dominant trace). See text for explanation.

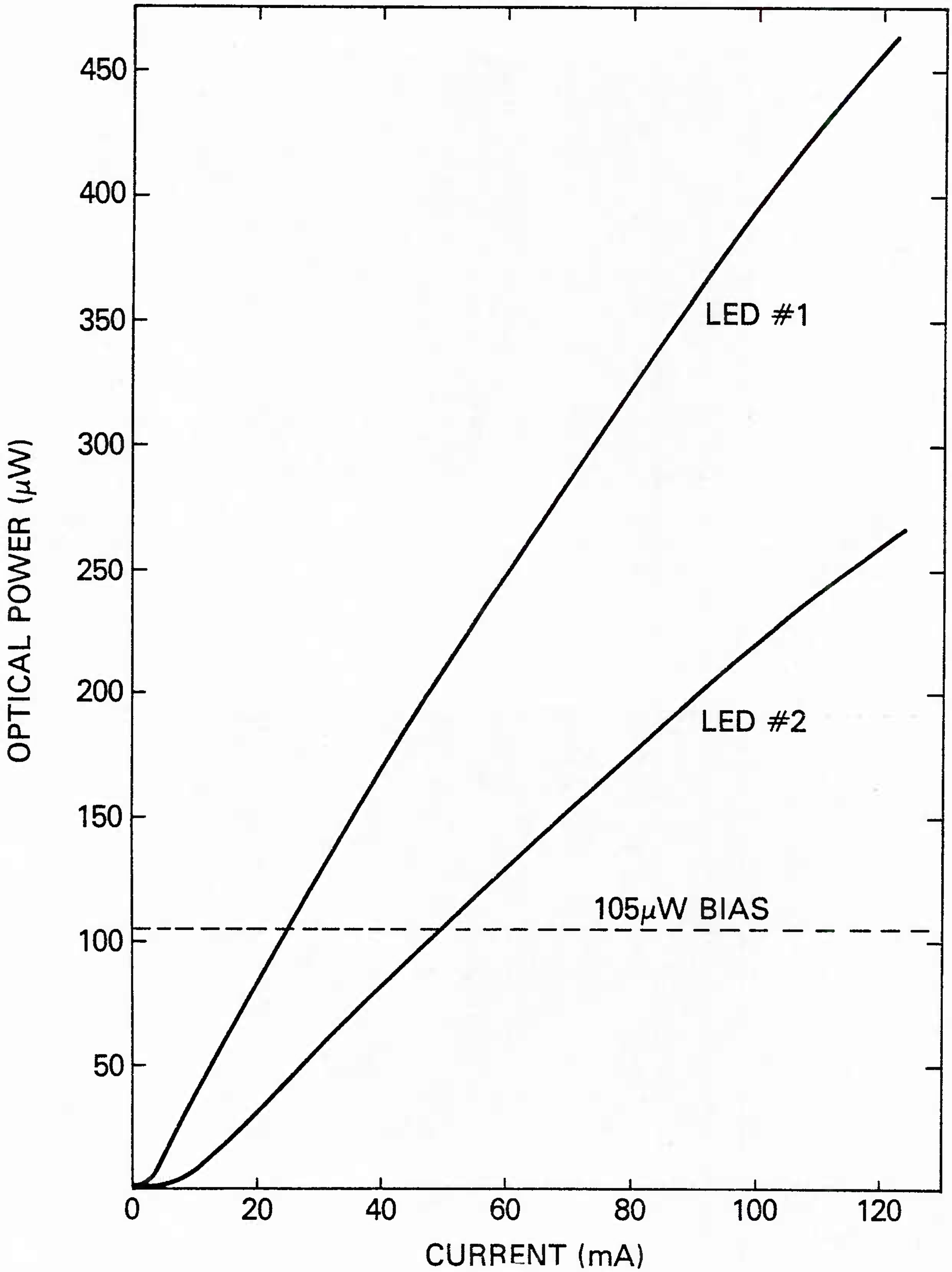


Fig. 13 — Optical transfer characteristics for the two LEDs.

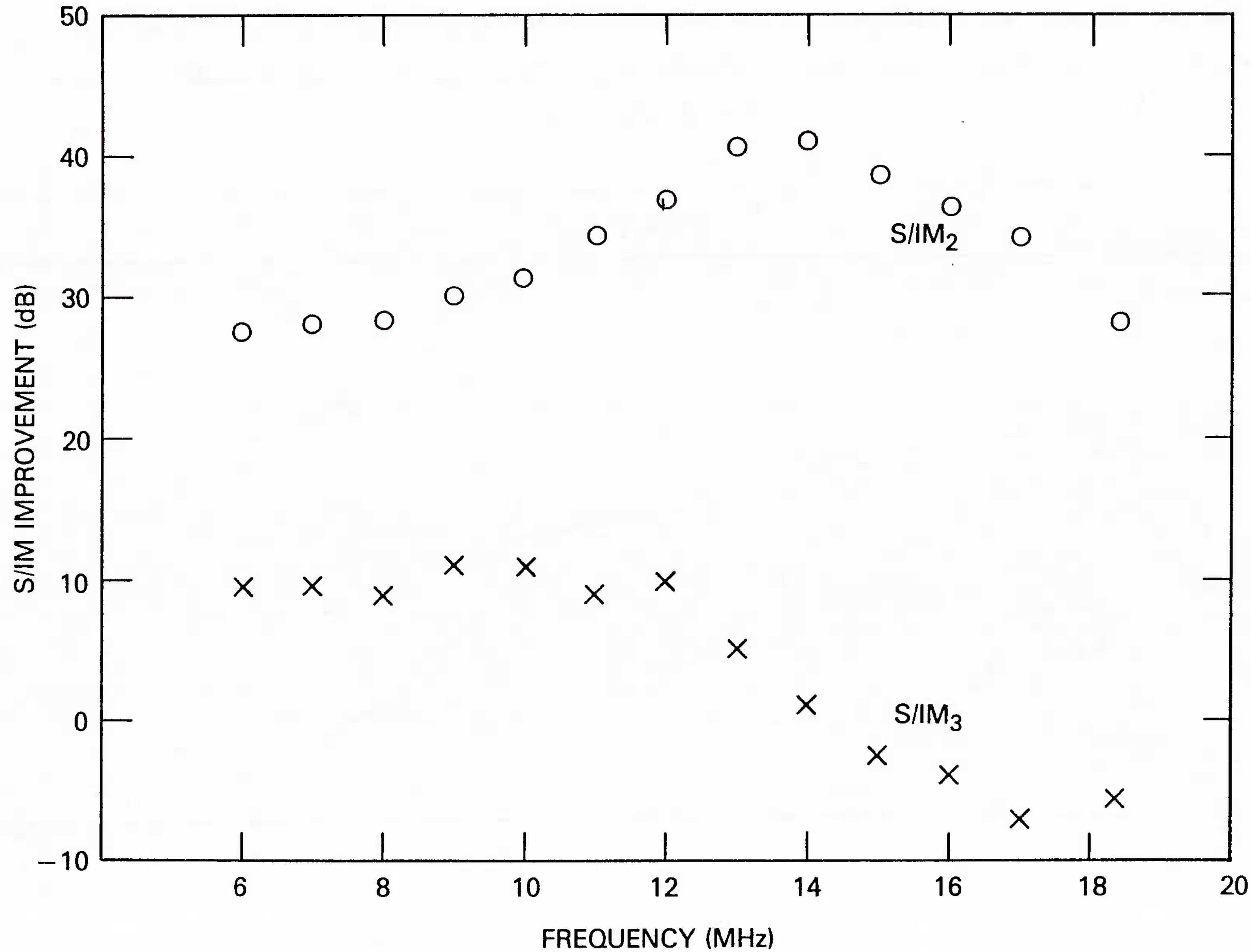


Fig. 14 — Improvement of the ratio of signal to intermodulation distortion resulting from quasi-feedforward linearization.

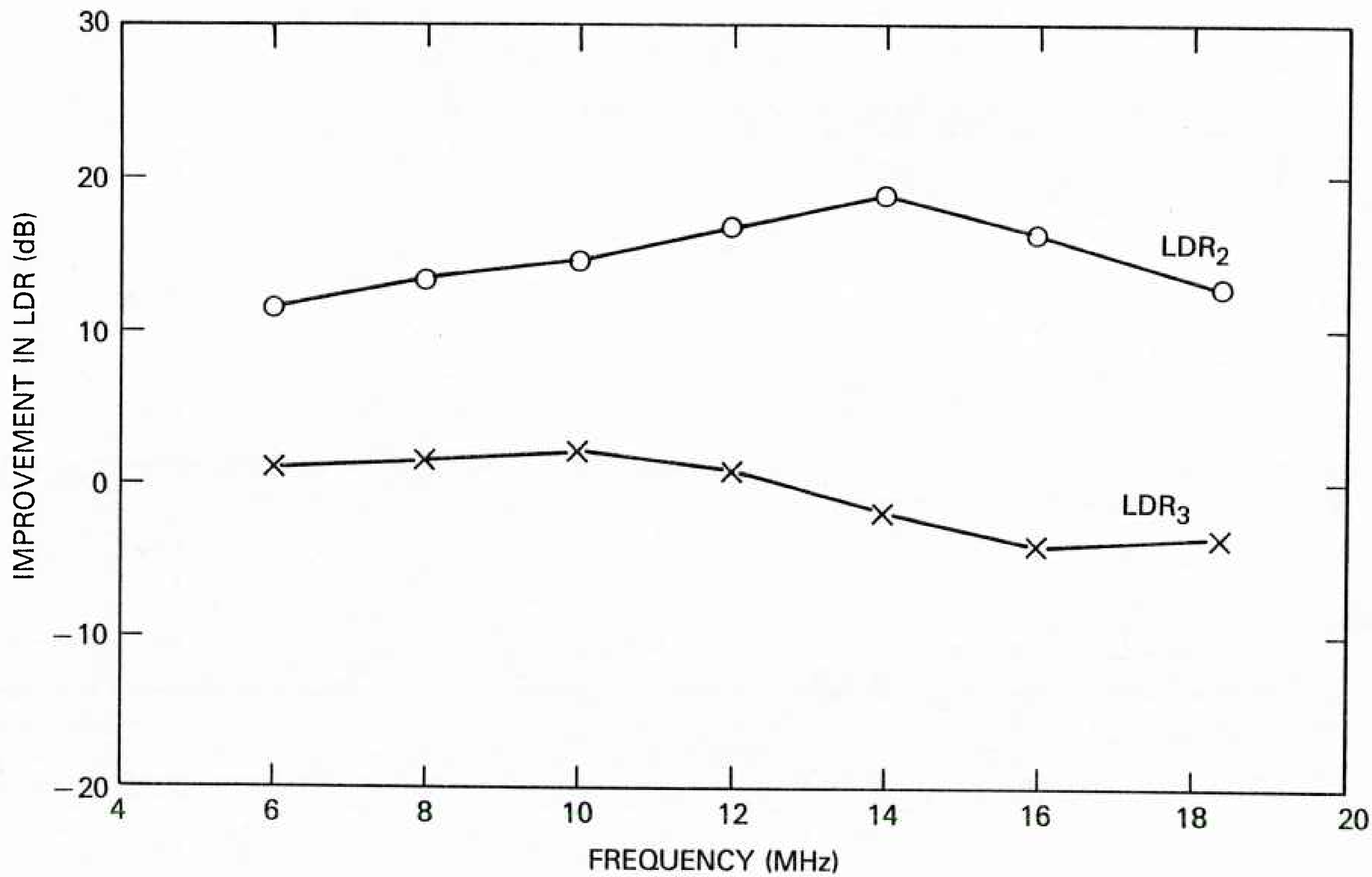


Fig. 15 — Improvement in linear dynamic range resulting from quasi-feedforward linearization.

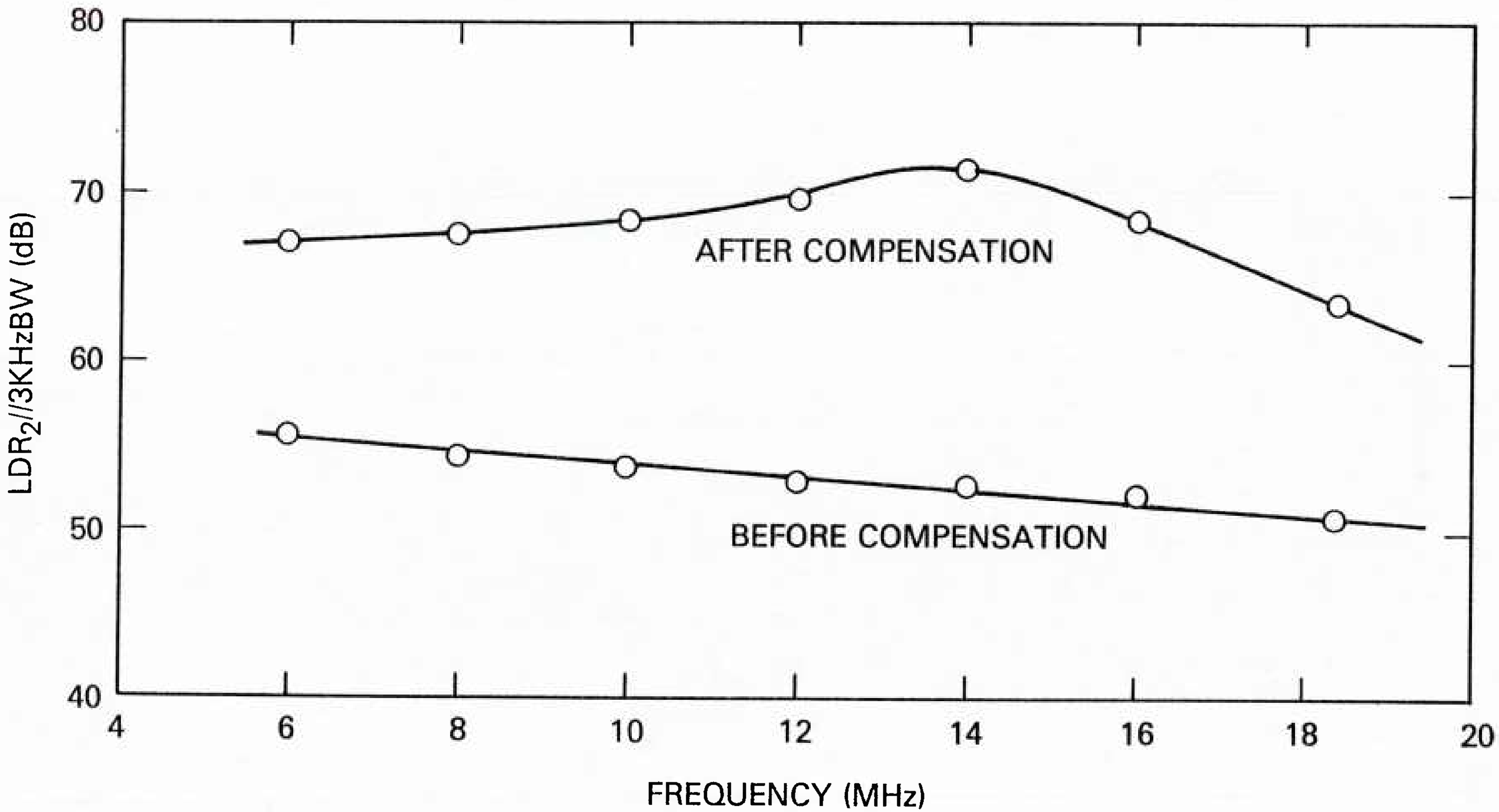


Fig. 16 — Second-order linear dynamic range versus frequency before and after linearization.

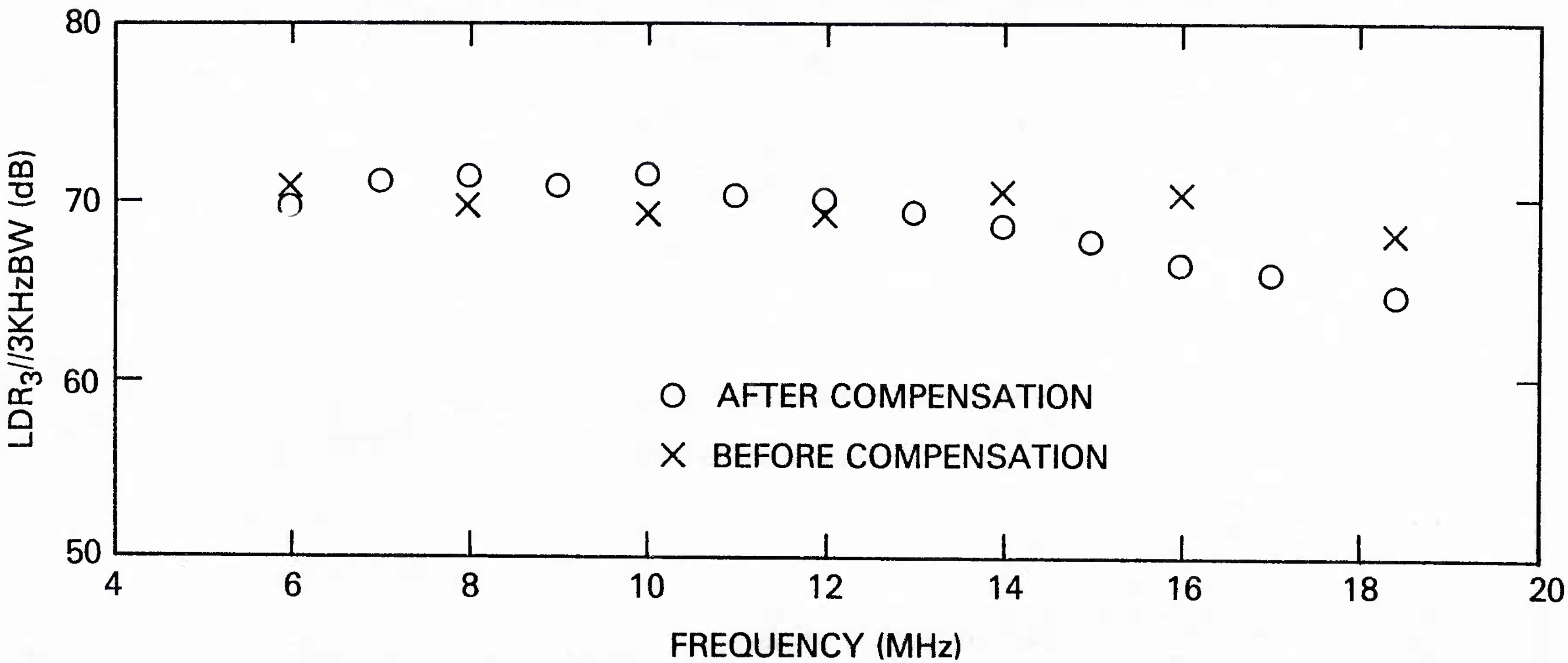


Fig. 17 — Third-order linear dynamic range versus frequency before and after linearization.

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